ACTIVE ELLIPTIC AUDIO FILTER DESIGN USING OP-AMPS

(Part 1)

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Introduction

Modern shortwave transceiver specifications have driven mixer and i.f. designs to near perfection in applying today's state-of-the-art technology. Nevertheless, the audio stages cannot claim similar advances, and in that respect seem to have been neglected. Perhaps one might argue that hi-fi quality is not needed because of the high atmospheric and man-made noise in this particular spectrum of radio frequencies, or that the extra cost of additional stages brings no meaningful improvement to transceiver specifications, tested under laboratory conditions. Certainly, from an operating point of view an optional state-of-the-art audio filter would be much appreciated, since it could make the difference between either working a station or losing it.

Very few published active filter designs actually describe an elliptic filter, mainly because of the overall complexity in designing and constructing one. Previous articles in *Radio Communication* [1, 2 and 3] inspired enthusiastic home-constructors to utilize inductors and capacitors to realize the elliptic transfer functions. This article is aimed as an introduction to active, fixed frequency audio filters being synthesized by a second method to achieve the elliptic filter characteristic, leading through the design stages and finishing with the construction of a narrow cw bandpass.

The technique employed uses RC active filter sections allowing latitude to the designer in choice of component values. Here, resistors and capacitors are used in conjunction with operational amplifiers to create a better, or at least the same, effect as can be obtained with a combination of passive elements.

The active approach offers many advantages. First, realizing that the inductive component presents the main difficulty in passive filter design. Inductors at audio frequencies tend to be bulky, heavy and expensive, and losses are often high. Non-standard inductance values necessitate winding by hand, which can prove a very cumbersome task.

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He joined Racal Communications UK in 1979, and currently heads the calibration laboratory of the data communications group, Racal Milgo at Warrington.

His amateur radio interests are concentrated on the home-construction of filters, direct-conversion third-method transceivers and data communication devices.

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Front view of the cw filter

Conversely, resistors and capacitors are compact and lightweight, also less expensive. The low-noise Texas Instruments TLO74 op-amp used in this design features all required parameters in a quad package, reducing the overall dimensions of the filter to a "handy" size. The op-amp high impedance jfet inputs allow lower-value capacitors to be used, and filter input and output impedance matching is easy, which all LC selective filters so critically depend upon.

The biggest advantage from the designer's point of view is the ability to precisely tune the active elliptic filter by trimming the value of certain resistors to obtain the desired calculated response, whereas LC passive filters are restricted to standard value components only, obtaining at best an approximated response.

Passive filters always introduce an insertion loss of some amount, whereas its active counterpart may exhibit gain, making it quite attractive for applications like the direct-conversion concept, where selectivity must originate at audio frequencies, and very high amplification is needed in a single section. Active filters require at least one single voltage rail, proving less of a constraint to the biquad elliptic type, since voltages between 6V and 36V at a few milliamps are easily tapped off existing equipment.

Virtually every transfer function may be performed, including the lowpass, highpass, bandpass, band-reject or notch, as well as the allpass, functions. For design and construction purposes perhaps the most interesting point to mention beforehand is that the basic building block of an active elliptic filter, a second-order stage (detailed in Fig 1(a)), always remains the same—only the RC components change in value to determine whether it becomes an lp, hp, bp, br or ap. One pcb layout caters for all types and possible combinations thereof.

The active elliptic lowpass filter will be considered first.

The biquad elliptic lowpass

A lowpass is a filter that passes signals with low frequencies and blocks those with high frequencies.

There are many types of lowpass filters; the four most popular being the Butterworth, Chebyshev, inverse Chebyshev, and Cauer or elliptic filters. The circuit of Fig 1(a) represents a biquad elliptic filter of second-order (N = 2) which is similar in complexity to a voltage-controlled voltage source (vcvs) elliptic or to a three-capacitor elliptic circuit. The biquad is easier to tune than a vcvs and has the advantage over the three-capacitor circuit that the gain may be set without having to trim one capacitor value. Also, as the name of the latter implies, it needs one additional capacitor.

The biquad gets its name from the input integrator or "area-under-the-curve device" using R4 as a damping resistor in parallel with capacitor C1. The second integrator with C2 does not have a damping resistor, and the remainder of the circuit is made up of a two-input, inverting current-summing amplifier. The gain (K) of this second-order stage is an inverting gain of -K. The elliptic or Cauer filter has an amplitude response showing ripples in both the pass- and stopbands, and a null just outside the passband. Each further second-order stage produces another zero in the stopband. These notches could, for example, be put where unwanted harmonics appear. The elliptic filter gives the fastest possible fall-off, in that for a given order (N) and allowable passband ripple width (prw) and minimum stopband loss (msl), it has the shortest transition width (tw).

An example of a fifth-order (N = 5) elliptic amplitude response is shown in Fig 2, where the following terminology applies:

(*a*) The transfer function of a realizable filter is a ratio of polynomials, where the degree n of the denominator polynomial is defined to be the "order" of the filter.

(*b*) The passband is the range of frequencies where the amplitude is greater than some specified number A1.

(*c*) The passband ripples are equal in magnitude and may be characterized by the maximum allowable deviation, which is called passband ripple width, prw = $-20 \log A1$ (dB).

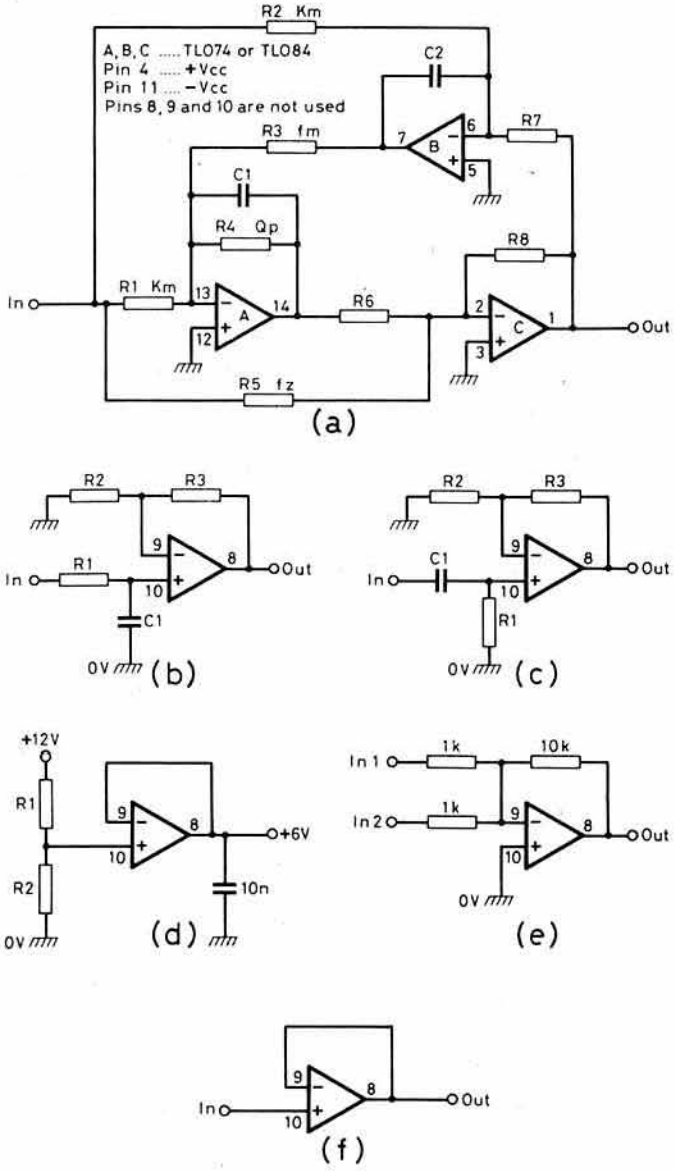


Fig 1. (a) Biquad elliptic second-order stage. (b) First-order lowpass. (c) First-order highpass. (d) Rail divider. (e) Gain-of-10 inverting summing amplifier. (f) Non-inverting unity gain buffer

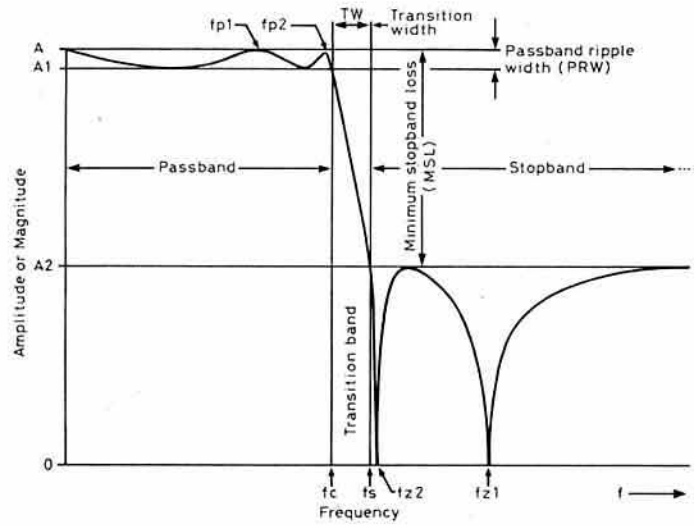
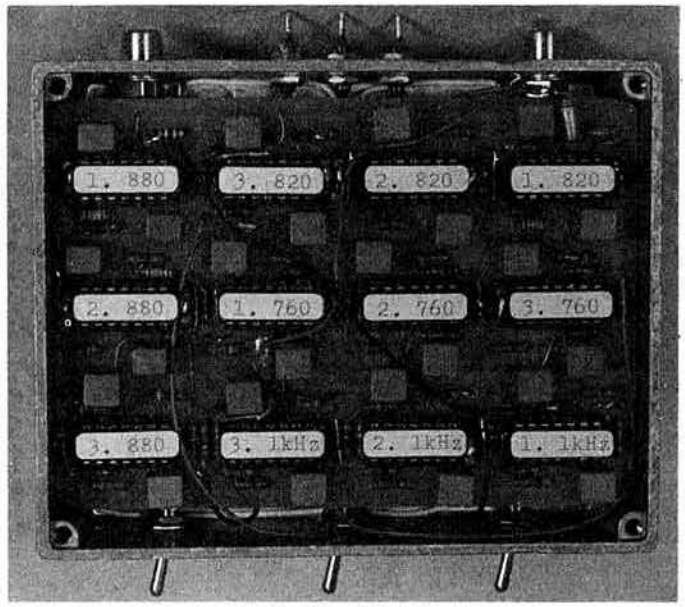


Fig 2. Elliptic lowpass response



Inside view of the filter

(*d*) The stopband is the range of frequencies where the amplitude is less than some specified amount A2.

(*e*) The stopband ripples are also equal in magnitude (although not necessarily equal to prw) and are characterized by the minimum stopband loss, msl = − 20 log A2 (dB).

(*f*) The transition band is the interval in which the response continually decreases from the passband to the stopband.

(*g*) The transition width is given by tw = fs − fc, as the band of frequencies in hertz, asserting the steepness of a filter. In its normalized form tw is given as a coefficient.

(*h*) The cut-off frequency is specified at the point of maximum passband ripple excursion, compared to other filter types, where the − 3dB-point characterizes the cut-off frequency. In the elliptic case a − 3dB-point only coincides if the passband ripple width is 3dB. (Elliptic filter example: prw = 0·5dB, then fc starts at − 0·5dB).

The amplitude or magnitude is usually expressed in decibels, where the prw cannot exceed 3dB and typical values for the msl or attenuation are considerably larger, possibly in the range between 20dB and 100dB, in which case we have the corresponding factors from 0·1 to 0·00001.

The optimum filter is the one whose amplitude response satisfies the following condition:

For a given order N and ripple prw and loss msl, to produce a minimum

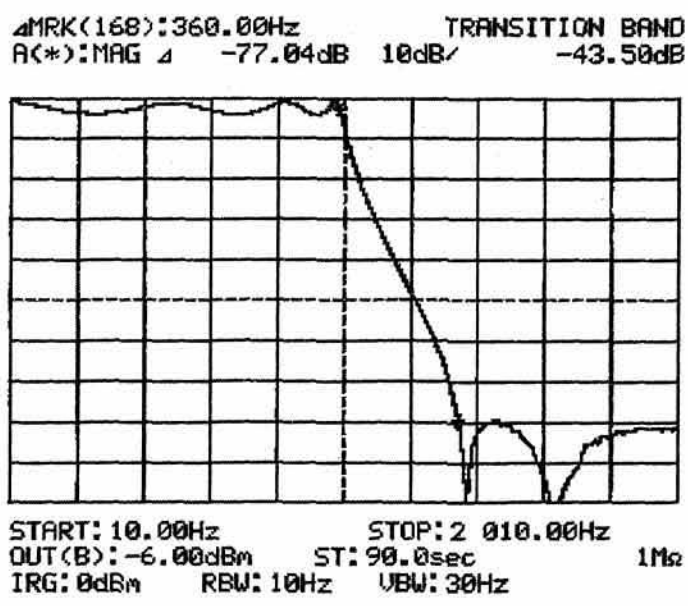


Fig 3. 1,000Hz lowpass response, transition width 360Hz. N = 7, prw = 3dB, msl = 80dB, fs = 1,360Hz

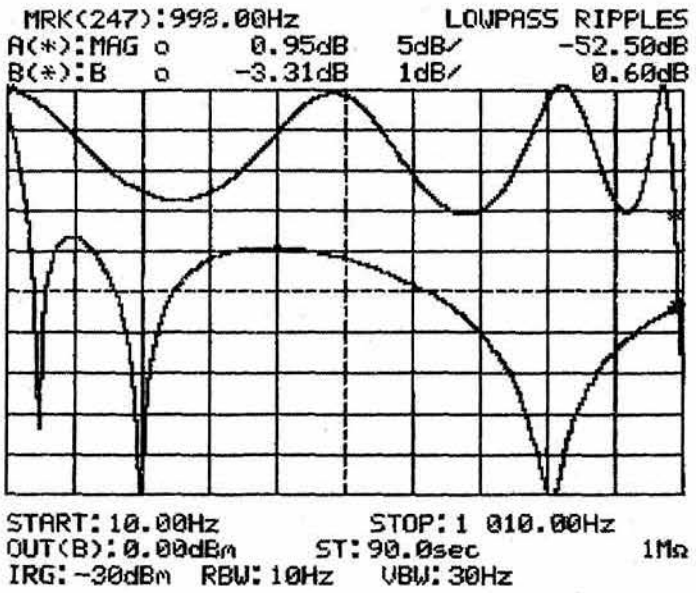


Fig 4. 1,000Hz lowpass ripple responses: top, 3dB passband ripple; bottom, >30dB stopband ripple

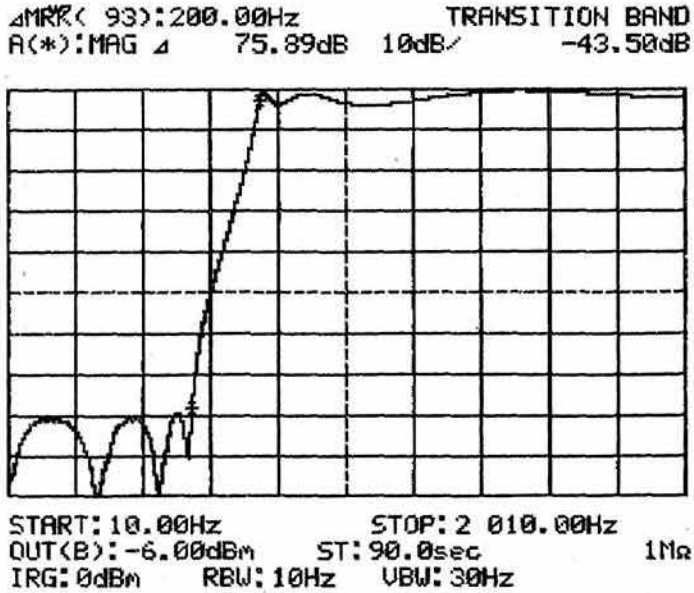


Fig 6. 760Hz highpass response, transition width 200Hz. N = 7, prw = 3dB, msl = 80dB, fs = 560Hz

transition width tw. That is, if A, A1, A2, N and the cut-off frequency fc are given, then the stopband frequency fs is a minimum.

In the all-pole case, the optimum filter is the Chebyshev. However, in the general case, the elliptic filter is optimum and far superior to the Chebyshev filter. For example, to match the transition with tw = 0·03 of a 12th-order elliptic filter, a Chebyshev filter with 39 orders would be required, as would a lot of compromising.

A typical lowpass response and its associated ripples are shown in Figs 3 and 4.

The biquad elliptic highpass

A highpass is a filter that passes signals with high frequencies and blocks those with low frequencies.

The transfer function of a highpass may be obtained from that of a lowpass filter using the process of mathematical transformation by 1/f, which will produce the mirror image of its lowpass counterpart. The information given and terminology used earlier in describing the biquad elliptic lowpass filter, also applies to the highpass, apart from some minor detail; ie the transition width in terms of frequency can be obtained by subtracting the smaller value stopband frequency from the larger cut-off frequency, or tw = fc-fs. Using tw as a coefficient, its frequency then becomes tw(hp) = tw/(1 + tw)*fc, in hertz. As in the lowpass, tw is a more preferred term for the elliptic filters, rather than decibels/octave, used in more shallow designs.

All the formula changes have been made, and we can use the result in a later chapter dealing with the calculation of the individual highpass elements.

Theoretically, a highpass should retain a flat passband response to infinite frequency, but in practice the active highpass has an upper frequency limit imposed on it, appearing as a second cut-off frequency, and normally determined by the gain/bandwidth product of the op-amp in use. A "natural" bandpass response is created. This effect can be seen in Fig 5, where the active highpass starts to roll off just below 1MHz, and reaches a minimum level (70dB down) at 4·6MHz. For audio applications this natural phenomenon leaves the frequency range of interest unaffected.

A typical highpass response and its associated ripples are shown in Figs 6 and 7.

The combination biquad elliptic bandpass

A bandpass is a filter that passes signals within a band of frequencies and blocks those below and above that band.

Normally, the bandwidth (bw) is given, which is centred approximately about the centre frequency fo. The actual centre frequency fo = SQR(fl*fu),

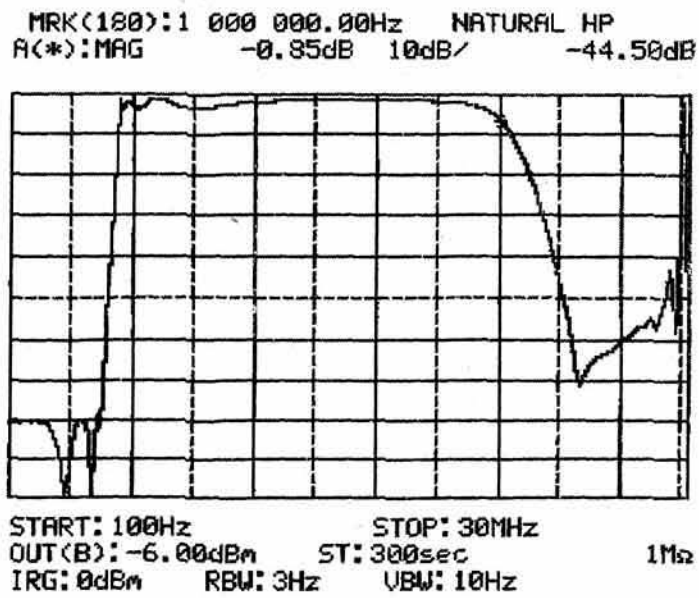


Fig 5. Active highpass natural response. 1MHz at −6dB, 4·6MHz at −70dB

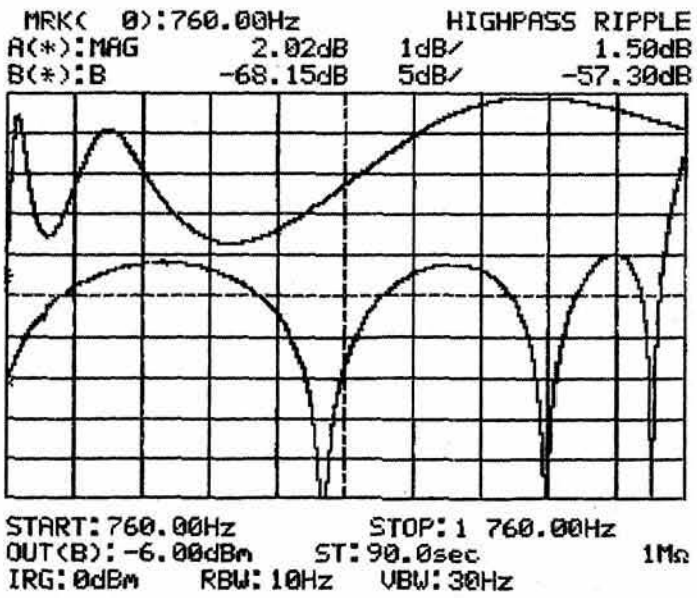


Fig 7. Highpass ripple responses: top, 3·6dB passband ripple; bottom, >30dB stopband ripple

where fl is the lower, and fu the upper cut-off frequency of the bandpass filter. The ratio Q = fo/bw is the quality factor of the bandpass and is a measure of its selectivity or sharpness. High values of Q correspond to a relative narrow bandwidth, and low values of Q to relative wide bandwidth —not to be mistaken for the Q of capacitors, as mentioned later, or the pole-pair quality factor Qp, as a measure of sensitivity to component tolerances for lowpass and highpass filters. As a rule of thumb, a true bandpass should be chosen if the percentage bandwidth is smaller than 75 per cent. A combination or overlapping bandpass, made up by cascading a low- and highpass, should be chosen if the percentage bandwidth is greater than 75 per cent.

As a practical example, suppose our centre frequency fo = 1,000Hz, and the bandwidth bw = 600Hz; the true bandpass should be chosen as pointed out above, since the percentage bandwidth is 600/1000*100 = 60 per cent, which is smaller than the rule of thumb figure of 75 per cent.

However, the calculations for a true biquad elliptic bandpass filter are rather more involved, and the tuning of its stages becomes more delicate, so that a separate article would have to be written to cover this topic in detail. Here we use the combination bandpass to great advantage, and aim for much smaller bandwidths than 600Hz, as can be seen later.

If we chose the highpass to be on the signal input side and the lowpass on the output, then capacitive coupling from the signal source to the first-order highpass stage will provide dc separation (see Fig 1(c)). Therefore, ground return resistors on the filter inputs are no longer needed, since each input stage will be driven by a low-impedance op-amp output. Putting the lowpass last has the added advantage of very low noise and harmonic distortion. The measured noise floor within the filter passband, when no signal is applied, was smaller than 80dB down against a reference level of 0dBm (0·775Vrms into 600Ω). This means that the internally-generated filter noise was smaller than 75μV.

A combined bandpass response is shown in Fig 8, where two narrow filter curves are superimposed on to the same graph, displaying steep slopes and very high stopband attenuation on a linear frequency scale. Fig 9 zooms in on the passband ripples of the two filters, where one vertical division on the graph represents 0·5dB. The top trace, with its wider bandwidth, has a maximum passband ripple excursion of 2·2dB, and the narrower trace below only totals 0·4dB. Originally, the biquad elliptic low- and highpass filters had a calculated prw of 3dB, but in cascading them, where the resulting ripple could have been anywhere between 0dB and 6dB, they complemented each other and produced a lower than calculated ripple response. This effect was a function of the chosen cut-off frequencies and cannot be found in the true bandpass design.

In order to compare the performance of various bandpass filters, a more descriptive term has been defined, called skirt selectivity or shape factor (rather than using Q = fo/bw). The shape factor of a bandpass is usually expressed as the bandwidth ratio at different attenuation levels, and is always greater than one. Preferred levels are the 30/3dB and 60/6dB bandwidths, but any other levels may be selected, as long as they remain the same for the purpose of comparing bandpass filters. See "measured results" for details on the shape factor filter performance figures.

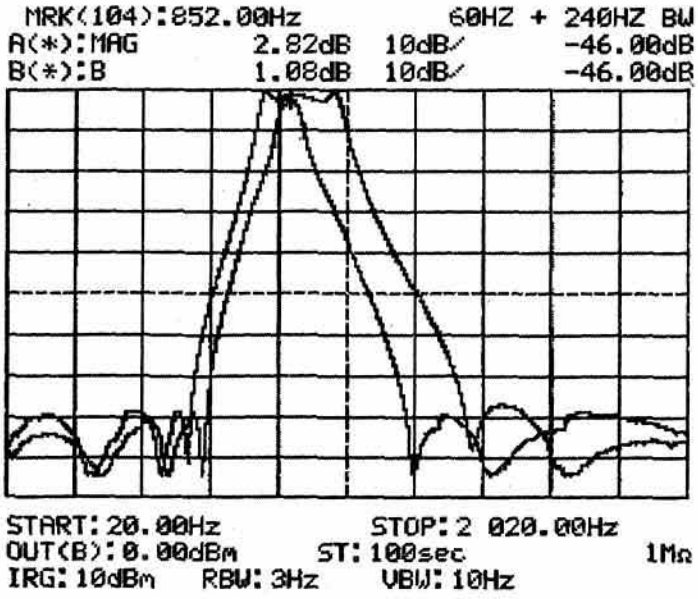


Fig 8. 60Hz and 240Hz bandpass responses (linear), cascaded lowpass and highpass filters. N = 7, prw = 3dB, msl = 80dB

Practical design considerations

To take a practical example for amateur radio use, let us assume the overall selectivity for cw in our receiver is lacking because the superhet was designed primarily for ssb with an i.f. bandwidth of approximately 2·4kHz, and its passband in the range from 300Hz to 2·7kHz. Since we are keen cw operators on the 7MHz amateur band, the need for an optional audio filter is immediately justified to reduce the effects of QRM.

Before quickly deciding on the final filter shape, let us look at some of the more important practical points.

The optimum listening frequency produced by received cw characters will greatly depend on the individual operator's hearing ability, and has been found to be a note or tone from as low as 400Hz to greater than 1,200Hz. It is also affected by QRM, which will push it to one or the other extremes, or by working a net with more than one station involved, each having a different pitch. In general, every cw operator will have his own preferred listening frequency, which should coincide with the transmitter's sidetone (normally adjustable). In my case that optimum frequency is about 850Hz. For reasons just mentioned above, the ideal design would be that which provides continuous variable tuning across a specified range of audio frequencies without changing the shape factor. This is possible with biquad elliptic filters using analogue switches being enabled by a variable duty-cycle generator. The solution in our case, dealing with fixed frequency types, is to provide switchable bandwidths using the combined biquad elliptic bandpass.

The first switch position should be a direct path, without any filters in circuit. This position is particularly useful for tuning, or monitoring of activities on the band. At the same time it can provide, for comparison, checks on the performance of any selected filters. To make a cw filter user-friendly, at least two different bandwidths should be available; a narrow one for normal cw traffic on a crowded band, and a very narrow one for when the going gets tough.

How narrow can we actually make the bandwidth of a cw filter? Theoretically, only a few hertz are needed for fast manually-sent morse, so a limit is imposed mainly by the technology utilized. Here the aim is to reproduce a clean note without any overshoot or ringing plus any other unwanted side-effects. On the other hand, it would be of little help to have a super-narrow filter, to find that the signal you were listening to has drifted outside the passband. Drift, therefore, is another factor, both your own and the other station's.

In the age of phase-locked loop synthesizers, where signals should remain stable over the full length of a cw QSO, a practical limit can be set below 100Hz. My first choice fell on 240Hz for the narrow bandwidth and 60Hz for the very narrow bw; primarily because when two combined bandpass filters are constructed from one highpass and one lowpass in each, two secondary bandwidths are created by cross-combination, as will be seen from the block diagram in Fig 13. It now remains to select the appropriate cut-off frequencies for each individual bandpass, bearing in mind that the preferred listening frequency and sidetone of 850Hz should be allowed to pass through any of them. The chosen frequencies are as follows:

	BW (Hz)	HP + LP, fc (Hz)
Primary	240	760-1,000
Secondary	180	820-1,000
Secondary	120	760-880
Primary	60	820-880

Switching is straightforward and has been arranged using single-pole changeover toggles, where the "direct" switch is positioned farthest left and the other two switches follow in line from the lowest cut-off frequency to the highest, as the furthest right toggle. Care should be taken when the actual connections are being made, since the wanted contact is usually in the opposite direction to the pointing-switch lever. The position of these switches and their marking can be seen on the photograph showing the front view of the cw filter, where each toggle points in the direction of its selected function.

Measurements have confirmed that the human ear is fairly tolerant to changes in amplitude within the passband, therefore we choose our prw to be the maximum possible, 3dB. Due to the high power density spectrum on 7MHz, 40dB of unwanted signal rejection is not going to do the job satisfactorily. Imagine you are listening to a weak station of −40dB, and a strong interfering station puts its carrier on the air at a level of 0dB; in other words, it is 100 times stronger. Should the interfering signal fall into the stopband of a 40dB filter, we could hear it at a level of −40dB, just the same as the wanted signal. Therefore, we should aim for rejection levels equal to or greater than 60dB. We select the msl at 80dB, which constitutes a compromise, because as the msl reaches its maximum at 100dB, the filter slope becomes less steep. Taking a closer look at the graphs in Figs 10 and 11 or the listed figures in Tables 1 and 2, where a set of parameters is offered for 3dB ripple filters, we can find out how complex, low down, steep and trimmable an envisaged filter is going to be.

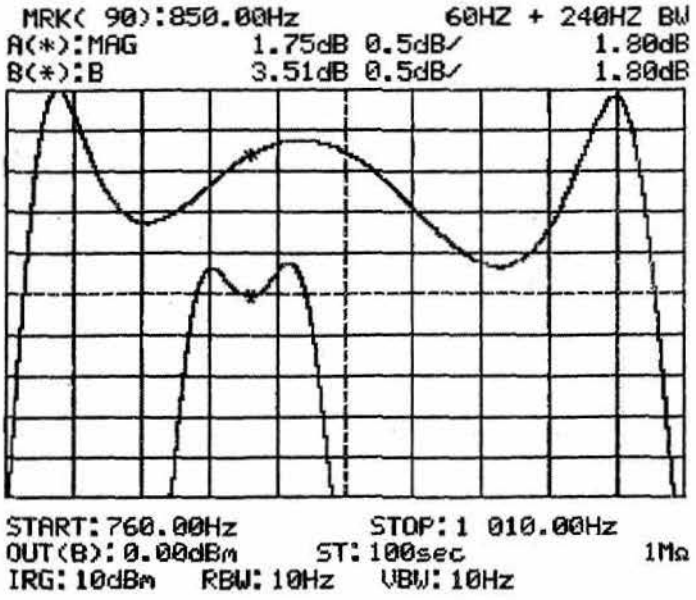


Fig 9. 60Hz and 240Hz bandpass ripple responses: top, bw = 240Hz, 2·2dB passband ripple; bottom, bw = 60Hz, 0·4dB passband ripple

Fig 10 and Table 1 give the starting point for 103 possible solutions if the problem had to be solved from a particular shape or steepness of the slope, whereas Fig 11 and Table 2 provide the same number of solutions if it had to be solved from a quality factor (Qp) point of view. The highest Qp is found in the last of the individually-cascaded second-order stages.

The normalized tw coefficient in Table 1 for an msl of 80dB can be read across, starting with 7·5859. A rather gradual roll-off would be experienced with only a third-order (N = 3) elliptic filter, and the stopband frequency (fs) will only reach its 80dB attenuation level at fs = (fc*tw) + fc = (1,000*7·5859) + 1,000 = 8,586Hz, an unacceptable shape in our case.

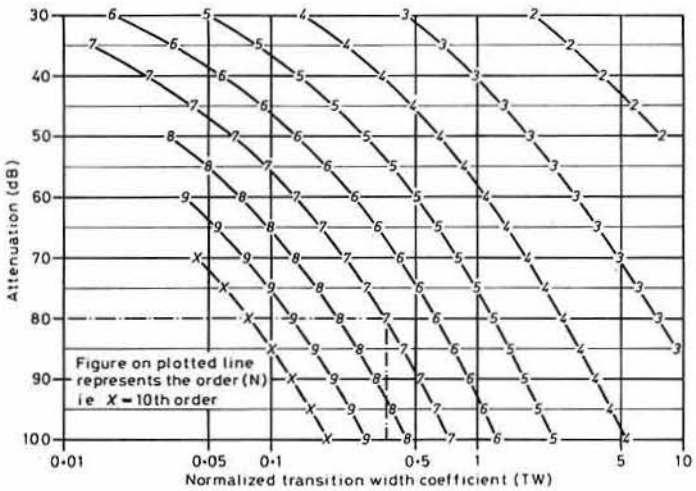

Attenuation (dB)
Figure on plotted line
represents the order (N)
ie X = 10th order
Normalized transition width coefficient (TW)

Fig 10. Attenuation versus tw. Transition width for elliptic filters with 3·0dB passband ripple (prw)

Table 1. Transition widths (tw) for elliptic filters with 3·0dB passband ripple (prw)

MSL	N=2	N=3	N=4	N=5	N=6	N=7	N=8	N=9	N=10
30	1·9032	0·4581	0·1454	0·0502	0·0178	—	—	—	—
35	2·8202	0·6878	0·2331	0·0878	0·0344	0·0137	—	—	—
40	4·0558	0·9802	0·3466	0·1393	0·0589	0·0254	—	—	—
45	5·7129	1·3460	0·4884	0·2058	0·0923	0·0425	—	—	—
50	7·9300	1·7986	0·6615	0·2885	0·1353	0·0657	0·0324	—	—
55	—	2·3547	0·8692	0·3885	0·1888	0·0955	0·0493	—	—
60	—	3·0347	1·1160	0·5071	0·2533	0·1325	0·0710	0·0386	—
65	—	3·8638	1·4067	0·6459	0·3294	0·1771	0·0980	0·0550	—
70	—	4·8725	1·7475	0·8068	0·4180	0·2297	0·1304	0·0754	0·0441
75	—	6·0981	2·1453	0·9918	0·5198	0·2908	0·1687	0·0999	0·0599
80	—	7·5859	2·6085	1·2032	0·6357	0·3607	0·2130	0·1288	0·0790
85	—	9·3907	3·1465	1·4440	0·7668	0·4399	0·2637	0·1623	0·1015
90	—	—	3·7706	1·7173	0·9143	0·5290	0·3210	0·2005	0·1275
95	—	—	4·4936	2·0267	1·0796	0·6285	0·3852	0·2437	0·1573
100	—	—	5·3307	2·3763	1·2640	0·7391	0·4568	0·2921	0·1908

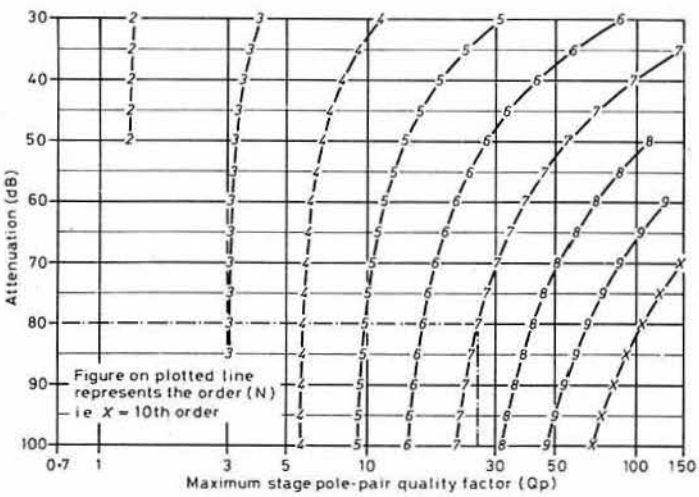


Fig 11. Attenuation versus Ap. Maximum space pole-pair quality factors (Qp) for elliptic filters with 3·0dB prw

Table 2. Max stage pole-pair quality factors (Qp) for elliptic filters with 3·0dB prw

MSL	N=2	N=3	N=4	N=5	N=6	N=7	N=8	N=9	N=10
30	1·3566	4·0264	11·268	31·264	86·646	—	—	—	—
35	1·3333	3·6811	9·3237	23·253	57·845	143·84	—	—	—
40	1·3206	3·4687	8·1408	18·653	42·541	96·928	—	—	—
45	1·3136	3·3335	7·3776	15·790	33·540	71·123	—	—	—
50	1·3097	3·2455	6·8649	13·901	27·838	55·588	110·92	—	—
55	—	3·1873	6·5101	12·600	24·018	45·586	86·421	—	—
60	—	3·1485	6·2594	11·675	21·349	38·801	70·394	127·63	—
65	—	3·1224	6·0796	11·001	19·421	34·007	59·388	103·62	—
70	—	3·1048	5·9491	10·500	17·992	30·508	51·537	86·950	146·62
75	—	3·0929	5·8536	10·123	16·911	27·885	45·757	74·944	122·67
80	—	3·0848	5·7833	9·8351	16·080	25·879	41·391	66·040	105·26
85	—	3·0793	5·7312	9·6142	15·432	24·316	38·023	59·267	92·258
90	—	—	5·6926	9·4432	14·922	23·084	35·380	54·010	82·303
95	—	—	5·6639	9·3100	14·517	22·098	33·273	49·856	74·539
100	—	—	5·6425	9·2060	14·193	21·303	31·576	46·528	68·367

To the far right of Table 1 we read 0·079 as the tw for the attenuation msl = 80dB. Going up on the vertical axis it shows that we need N = 10; in other words, five second-order stages to get the stopband frequency fs = 1,079Hz. A very near ideal shape, only 79Hz away and 80dB down —fantastic. Here the total stage count would amount to 4*5 = 20 second-order stages or, in terms of components, a total of 20 ics plus passive components—rather complex as an introductory project.

Also, if we look at Table 2 under N = 10 and across to msl = 80dB, we find that in this case the fifth second-order stage has the extremely high pole-pair quality factor (Qp) of greater than 105. Because high values of Qp will have a large spread of element values and become more sensitive to component tolerances, these stages would not escape elaborate tuning and therefore are not recommended for this project. As a rough guide, consider the following figures for Qp as an indication of how much trimming or tuning might be necessary in order to bring the constructed filter into agreement with its calculated response.

Qp	Component tolerances	Tuning or trimming requirements
0–10	R 5% and C 5%	Practically none
10–40	R 1% and C 2·5% or 5%	Very little
40–100	R 1%, measure C	Necessary
100–150	R 1%, measure C	Elaborate

The instruments required for the tuning process would consist of a set-up including: audio generator, oscilloscope and frequency counter.

If we search for Qp to be in a safer region, we find a seventh-order filter (N = 7) at 80dB down will have a pole-pair quality factor of 25·879, say 26. The accompanying transition width factor for N = 7 is given as tw = 0·3607 and will put the stopband frequency at fs = (fc*tw) + fc = (1,000*0·3607) + 1,000 = 1,360·7Hz; a much better practical solution to our problem, since N = 7 only needs three ics, and the total ic count falls to a reasonable figure of 4*3 = 12.

Odd-order filters (ie N = 3,5,7, or 9) offer a better performance for less components, and for this reason are more popular than even-order filters. Also, less calculations will be asked for. To build a seventh-order filter we simply cascade three stages of second-order as in Fig 1(a), and add one single or first-order stage, see Figs 1(b), 1(c), and 12.

The audio output from either a headphone or loudspeaker terminal already has sufficient amplitude, so no further gain from the filter is required. Therefore, the overall filter gain (K) is chosen to be one or unity.

So far we have gathered the following design parameters:
fc = 1,000Hz, as the LP1 cut-off frequency at minus 3dB;

fs = 1,361Hz, as the LP1 stopband frequency at minus 80dB;
fc = 880Hz, as the LP2 cut-off frequency at minus 3dB;
fs = 1,197Hz, as the LP2 stopband frequency at minus 80dB;
fc = 820Hz, as the HP2 cut-off frequency at minus 3dB;
fs = 603Hz, as the HP2 stopband frequency at minus 80dB;
fc = 760Hz, as the HP1 cut-off frequency at minus 3dB;
fs = 559Hz, as the HP1 stopband frequency at minus 80dB;
N = 7, seven orders are required, three second-order, plus one first-order;
prw = 3dB, maximum selected passband ripple width;
msl = 80dB, minimum stopband loss or attenuation;
K = 1 or 0dB, unity filter gain (same out as in);
tw = 0·3607, normalized transition width factor, common to all filters selected.

Normalized lowpass filter data and calculations

The mathematical process to arrive at an elliptic filter response requires the use of Jacobi elliptic functions [6], which are extremely tedious to calculate. For convenience, the normalized case (wc = 1 radian/s) of the transfer functions in factored form is tabulated as lowpass data for the orders n = N = 2 to 10. See extracted lowpass data in Table 3. (*To follow in Part 2.*) Available passband ripples are prw = 0·1, 0·5, 1, 2 and 3dB, and the msl, in most cases, is 30dB to 100dB in steps of 5dB.

TW is the normalized transition width which results, and A, B and C are the actual lowpass filter coefficients or second-order factors, where the first line or row of data corresponds to the first second-order stage or factor, the second line to the second factor, and so on. In the case of an odd-order filter (ie N = 7), the first-order factor is given in the last line of data under C.

The next five columns contain valuable information for component selection and the precise tuning or matching of the individual responses. Presented are the pole-pair frequency coefficient (Fp), the pole-pair quality factor (Qp), the stopband zero frequency coefficient (Fz) and, in the case of passband peaking, the peak frequency coefficient (Fm) and its appropriate level (Km) at which it occurs. If the peak frequency (fm) is an imaginary number, then there is no passband peaking.

Altogether there are 545 listed filter solutions, containing 33 pages of lowpass data like the one in Table 3, from which any elliptic filter may be calculated to fit a particular response, including hp, bp and notch filters. I can supply lowpass data printouts for a small charge to costs.

From our given design parameters we can now find the appropriate coefficients A, B, C and Fp (look up Table 3, under N = 7, prw = 3, msl = 80).

There are 10 individual filter elements or component values to be found for each second-order stage, four of which are arbitrarily chosen (C1, C2, R6 and R8) and six resistor values (R1 to R5, and R7), where the formula below apply. The calculations will have to be repeated three times (N = 7 has three second-order stages plus one first-order stage). Again the process seems a tedious one, and you may want to skip this part and have a look at Table 4, where all component values, design parameters, lowpass coefficients, tuning data and comments are presented as a computer printout for the 1kHz lowpass. Table 5 summarizes the 760Hz highpass results in detail. (*Tables 4 and 5 to follow in Part 2.*)

In view of the intricate complexity of these calculations (minimum maths), I am prepared to provide printed results like those in Tables 4 and 5 for readers who cannot get involved. A suggested format of info required for a customized active elliptic audio filter design can be found in the appendix, from which these printed results can be generated.

Biquad elliptic lowpass calculations

Calculate each second-order lowpass stage in turn, as per the following step-by-step example: LP fc = 1,000Hz.

(1) Find the circuit's angular frequency wc = W = 2*pi*fc = 6,283·18Hz (for ease of repetitive use).

(2) Find the pole-pair frequency fp = Fp*fc = 0·52445*1,000 = 524·45Hz.

(3) Find the stage zero frequency fz = Fz*fc = 1·626828*1,000 = 1,627Hz (rounded figure for tuning purposes).

(4) Select a standard value for capacitors C1 and C2. For moderate K and Qp let C1 = C2, preferably near 1E4/fp. C1 = C2 = 10,000/524·45 = 19·07nF.
Selected C1 = C2 = 10nF for all stages, because of size and availability.

(5) Select a standard value for resistors R6 and R8, preferably near 1/(W*C1).
R6 = R8 = 1/(6,283·18*10*1E-9) = 15,915Ω.
Selected R6 = R8 = 20kΩ for all stages, because this choice brings all other resistors very close to preferred standard values.

(6) R1 = A/(K*B*C*W*C1).
R1 = 2·646568/(1*0·244509*0·275048* 6,283·18*10*1E-9) = 626·33kΩ.

(7) R2 = 1/(K*Fp*W*C2).
R2 = 1/(1*0·52445*6,283·18*10*1E-9) = 30·35kΩ.

(8) R3 = 1/(Fp*W*C1).
R3 = 1/0·52445*6,283·18*10*1E-9) = 30·35kΩ.

(9) R4 = 1/(B*W*C1).
R4 = 1/(0·244509*6,283·18*10*1E-9) = 65·09kΩ.

(10) R5 = (A*R6)/(K*C).
R5 = (2·646568*20,000)/(1*0·275048) = 192·44kΩ.

(11) R7 = (C1*R3)/C2.
R7 = (10*1E-9*30,350)/10*1E-9 = 30·35kΩ.

Repeat calculations for the second and third stages, starting at (2), since fc remains the same. The formulas above are used for all biquad elliptic lowpass filter calculations.

The first-order stage in our case has unity gain, therefore the inverting input to the selected op-amp can simply be linked to the output of the same, (pins 8 and 9, see Fig 1(b)), where C1 is selected to be 10nF.

(12) R1 = 1/(W*C1*C) = 1/(6,283·18*10*1E-9*0·152661) = 104·25kΩ.
R2 and R3 are only necessary, should more than unity gain (K > 1) be asked for.

(13) R2 = (K*R1)/(K-1).

(14) R3 = K*R1.

A first-order stage only produces a pole and no zeros, where the frequency is:

(15) fp = Fp*fc = 0·390719*1,000 = 391Hz.

Select standard value resistors and capacitors as close as possible to the calculated values.

TO BE CONCLUDED

A DX TRAILER

David A Reid, GM0BZF*

WHEN I DECIDED to build this trailer, I thought that it was going to be a difficult and painstakingly expensive job. But I was wrong, it has turned out to be an easy (or relatively so) and enjoyable experience.

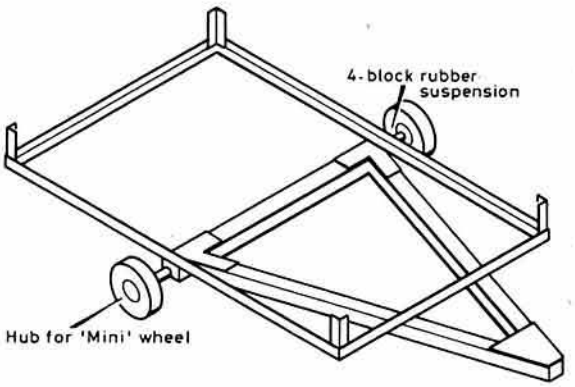


Fig 1. The basic chassis

David Reid was licensed as GM6JLQ in 1982, and became GM0BZF in June 1985. From 144MHz dxing in 1982, his interests now cover the range 3·5MHz dx to 1·3GHz—rtty on 144MHz; cw on all bands, receiving on 50MHz, and transmitting on 3·5, 14, 21, 28, 70 and 144MHz. He received the G-QRP Club award as the first Class B to work 50 stations with his variance as GM6JLQ. He is a member of the Clyde Coast Raynet Group; and the Helensburgh ARC, of which he is secretary. He is a microelectronics engineer by profession and is very interested in computers and radio interfacing.

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ACTIVE ELLIPTIC AUDIO FILTER DESIGN USING OP-AMPS

(Part 2)

D H G Fritsch, G0CKZ*

Biquad elliptic highpass calculations

Calculate each second-order highpass stage in turn, as per the following step-by-step example: HP fc = 760Hz.

(1) Find the circuit's angular frequency wc = W = 2*PI*fc = 4,775·22Hz (for ease of repetitive use).

(2) Find the pole-pair frequency fp = (1/Fp)*fc = (1/0·52445)*760 = 1,449·14Hz.

(3) Find the stage zero frequency fz = (1/Fz)*fc = (1/1·626828)*760 = 467Hz (rounded figure for tuning purposes).

(4) Select a standard value for capacitors C1 and C2.
For moderate K and Qp let C1 = C2, preferably near 1E4/f.
C1 = C2 = 10,000/1,449·14 = 6·9nF.
Selected C1 = C2 = 10nF for all stages, because of size and availability.

(5) Select a standard value for resistors R6 and R8, preferably near 1/(W*C1).
R6 = R8 = 1/(4,775·22*10*1E-9) = 20,941Ω.
Selected R6 = R8 = 20kΩ for all stages.

(6) R1 = C/(K*B*W*C1).
R1 = 0·275048/(1*0·244509*4,775·22*10*1E-9) = 23·56kΩ.

(7) R2 = A/(K*Fp*W*C2).
R2 = 2·646568/(1*0·52445*4,775·22*10*1E-9) = 105·68kΩ.

(8) R3 = Fp/(W*C1).
R3 = 0·52445/(4,775·22*10*1E-9) = 10·98kΩ.

(9) R4 = K*R1.
R4 = 1*23·56 = 23·56kΩ.

(10) R5 = R6/K.
R5 = 20/1 = 20kΩ.

(11) R7 = (C1*R3)/C2.
R7 = (10*1E-9*10,980)/10*1E-9 = 10·98kΩ.

Repeat calculations for the second and third stages, starting at (2), since fc remains the same. The formulas above are used for all biquad elliptic highpass filter calculations.

The first-order stage in our case has unity gain, therefore the inverting input to the selected op-amp can simply be linked to the output of the same, (pins 8 and 9, see Fig 1(c)), where C1 = 10nF.

(12) R1 = C/(W*C1) = 0·152661/(4,775·22*10*1E-9) = 3·2kΩ.

In this case the following values were selected new: R1 = 27kΩ and C1 = 1·2nF, to increase the stage input impedance, and for the convenience of using standard value components.

R2 and R3 are only necessary, should more than unity gain (K > 1) be asked for.

(13) R2 = (K*R1)/(K-1).

(14) R3 = K*R1.

A first-order stage only produces one pole and no zero. Its frequency is:

(15) fp = (1/FP)*fc = (1/0·390719)*760 = 1,945Hz.

Select standard value resistors and capacitors as close as possible to the calculated values.

Construction and tuning of the stages

The sensitivity of an active filter is a measure of how accurate its op-amp, resistor and capacitor tolerances have to be to get the response within certain limits of the one desired. For best results the choice of component elements would have to be considered, especially so if one of the four following terms or a combination thereof becomes high: cut-off frequency, gain, quality factor and order.

From the list of well-known op-amp manufacturers, ie Fairchild Semiconductor, National Semiconductor, Motorola, RCA, Signetics Corporation and Texas Instruments, the choice of op-amps is plentiful and the data sheets should provide adequate information about parameters, including open loop voltage gain, input resistance, input offset voltage, input offset current, input bias current, slew rate, drift with temperature coefficients, full power bandwidth etc.

The dc offset voltage and its drift at the op-amp output become important when a large number of stages are cascaded in higher-order filters. Output noise from the filter is due to the internal voltage noise and current noise generated by each individual op-amp. Current noise may cause greater noise outputs than voltage noise if the op-amp currents are flowing through large resistances. Fet op-amps have very low bias currents and also low current noise. The slew rate of an op-amp represents the limiting value of output voltage swing at a given frequency and is usually a number given in V/μs. The open loop gain is neither infinite nor constant for all frequencies and is given as the figure of full power bandwidth.

Since each second-order stage requires three single op-amps, the choice falls to a quad package for reasons of size and drift being contained on one substrate. Here the Texas Instrument bifet TL074 has been chosen for its superior performance, which is pin compatible to the TL084 of the same family with slightly inferior noise performance. The low-power TL064 and single-supply version TL094 can also be used in less critical applications, where slew rate and full power bandwidth are traded off against lower power consumption and single-rail applications, respectively. National Semiconductor types LM124, LM224, LM324 and LM348 are also pin compatible and may be used where the slew rate and input resistance are not of prime importance.

Other devices are the MC3403 from Motorola, the HA4741 or HA5144 from Harris, and so on, leaving plenty of scope for substitution or replacement.

Resistors present the least problem in filter design. Carbon composition types of 5 or 10 per cent tolerance should only be used in low sensitivity circuits. Metal-film resistors having a 1 or 2 per cent tolerance, 50ppm/°C temperature coefficient and lower noise figures should be preferred for high-performance applications.

In the case of capacitors, mica, polystyrene and polypropylene types offer properties of high stability, high insulation resistance, small temperature drifts and power factors. The merit of a capacitor from the point of freedom from losses is expressed in terms of the power factor. The power factor is the sine of the angle by which the current flowing into the capacitor fails to be 90° out of phase with the applied voltage. The tangent of this angle is called the dissipation factor. The reciprocal of the dissipation factor is called the Q, and is the ratio of the capacitor reactance to the equivalent series resistance.

Silvered mica capacitors offer the best temperature drift coefficients and tolerances with an inferior power factor, compared to the other two types, but their size and cost make them unsuitable. Polystyrene foil capacitors made by Suflex are available in 2·5 per cent tolerances with values up to 10nF, but with a slightly higher power factor than polypropylene types.

For practical reasons it is well worth spending a moment or two studying capacitor specifications before a choice is made, since capacitors present a more severe problem to high-sensitivity filter design (large Qp).

Best overall results were obtained using the Fkp2 series, a polypropylene film or foil constructed capacitor with a tolerance of five per cent, made by Wima. Its temperature coefficient is specified as ±100ppm/°C and its power factor smaller than 0·0004 at 1kHz, together with a very high insulation resistance of greater than 5*5*1E10MΩ. The working voltage is 63V, and it comes in a very small size of only 7·2 by 4·5 by 6mm (L,W,H) for values from 220pF to 2·2nF, with a standard pcb lead spacing of 5mm. The body colour is green and the largest value available from Farnell in Leeds is 10nF. Capacitors with five per cent, or even 2·5 per cent tolerance limits open avenues for errors in higher sensitivity circuits, but can still be put to very good use if their values are measured and then inserted as the formula values for capacitors C1 and C2.

The selected components used for the described lowpass and highpass filters were of the following preferred standard values:

LP 1,000Hz	1-stage	2-stage	3-stage
C1 and C2	10nF	10nF	10nF
R1	620kΩ	1M + 200kΩ in series	680kΩ + 110kΩ in series
R2, R3, R7	30kΩ	18kΩ	16kΩ
R4	62kΩ	120kΩ	390kΩ + 15kΩ in series
R5	200kΩ	200kΩ	39kΩ
R6 and R8	20kΩ	20kΩ	20kΩ

First-order stage C1 = 10nF and R1 = 100kΩ.

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Table 3. Normalized elliptic lowpass filter data, continued . . .

N	PRW	MSL	TW	A	B	C	Fp	Qp	Fz	Fm	Km
7	3.0	40	0.0254	1.190991	0.252082	0.513589	0.716651	2.8429	1.091325	0.661226	1.773
				2.238663	0.062408	0.896966	0.947083	15.1757	1.496216	0.944681	9.122
				1.061054	0.010289	0.994598	0.997295	96.9283	1.030075	0.996475	6.148
				*	*	0.238606	0.488473	*	*	*	*
7	3.0	45	0.0425	1.274060	0.255789	0.460612	0.678684	2.6533	1.128743	0.627333	1.830
				2.579873	0.074990	0.868304	0.931828	12.4260	1.606198	0.928789	8.271
				1.100306	0.014007	0.992457	0.996221	71.1231	1.048955	0.995266	7.086
				*	*	0.218823	0.467785	*	*	*	*
7	3.0	50	0.0657	1.377775	0.256657	0.416609	0.645453	2.5148	1.173787	0.597627	1.876
				2.979224	0.086497	0.840331	0.916696	10.5980	1.726048	0.913051	7.636
				1.153728	0.017901	0.990174	0.995075	55.5877	1.074117	0.994020	7.935
				*	*	0.202913	0.450459	*	*	*	*
7	3.0	55	0.0955	1.505082	0.255807	0.380296	0.616681	2.4107	1.226818	0.571842	1.912
				3.446897	0.096765	0.813967	0.902201	9.3236	1.856582	0.898000	7.149
				1.223507	0.021808	0.987853	0.993908	45.5858	1.106123	0.992786	8.326
				*	*	0.189998	0.435888	*	*	*	*
7	3.0	60	0.1325	1.659515	0.253994	0.350411	0.591955	2.3306	1.288222	0.549615	1.940
				3.995157	0.105768	0.789748	0.888678	8.4021	1.998789	0.883976	6.768
				1.312029	0.025536	0.985577	0.992762	38.8010	1.145488	0.991608	9.694
				*	*	0.179438	0.423601	*	*	*	*
7	3.0	65	0.1771	1.845308	0.251707	0.325825	0.570811	2.2678	1.358421	0.530552	1.962
				4.638605	0.113567	0.767917	0.876309	7.7162	2.153742	0.871168	6.465
				1.422015	0.029161	0.983407	0.991669	34.0067	1.192483	0.990494	10.523
				*	*	0.170752	0.413221	*	*	*	*
7	3.0	70	0.2297	2.067529	0.249253	0.305577	0.552790	2.2178	1.437890	0.514260	1.980
				5.394521	0.120266	0.748512	0.865166	7.1938	2.322611	0.859681	6.222
				1.556654	0.032472	0.981381	0.990647	30.5077	1.247659	0.989473	11.304
				*	*	0.163574	0.404443	*	*	*	*
7	3.0	75	0.2908	2.332220	0.246818	0.288368	0.537464	2.1776	1.527161	0.500370	1.993
				6.283299	0.125986	0.731445	0.855246	6.7884	2.506651	0.849371	6.024
				1.719727	0.035492	0.979522	0.989708	27.8854	1.311384	0.988548	12.028
				*	*	0.157618	0.397011	*	*	*	*
7	3.0	80	0.3607	2.646568	0.244509	0.275048	0.524450	2.1449	1.626828	0.488549	2.004
				7.328972	0.130851	0.716553	0.846495	6.4691	2.707207	0.840827	5.263
				1.915736	0.038211	0.977841	0.988858	25.8789	1.384101	0.987720	12.692
				*	*	0.152661	0.390719	*	*	*	*
7	3.0	85	0.4399	3.019103	0.242382	0.263589	0.513409	2.1182	1.737557	0.478502	2.013
				8.559849	0.134979	0.708637	0.838831	6.2145	2.925722	0.832411	5.729
				2.150047	0.040635	0.976335	0.988097	24.3164	1.466304	0.986984	13.294
				*	*	0.148522	0.385386	*	*	*	*
7	3.0	90	0.5290	3.459928	0.240462	0.254063	0.504047	2.0962	1.860088	0.469967	2.019
				10.009279	0.138475	0.692487	0.832158	6.0094	3.163744	0.825521	5.619
				2.429044	0.042776	0.975000	0.987421	23.0835	1.558539	0.986336	13.836
				*	*	0.145059	0.380866	*	*	*	*
7	3.0	95	0.6235	3.980998	0.238752	0.246126	0.496111	2.0779	1.995244	0.462728	2.025
				11.716544	0.141434	0.682895	0.826375	5.8428	3.422944	0.819552	5.528
				2.760309	0.044656	0.973825	0.986826	22.0984	1.661418	0.985770	14.318
				*	*	0.142155	0.377084	*	*	*	*
7	3.0	100	0.7391	4.596441	0.237245	0.239498	0.489385	2.0628	2.143931	0.456576	2.029
				13.727921	0.143936	0.674667	0.821381	5.7066	3.705121	0.814399	5.452
				3.152835	0.046298	0.972796	0.986304	21.3084	1.775622	0.985276	14.745
				*	*	0.139717	0.373787	*	*	*	*
8	0.1	30	0.0227	1.052604	1.152436	0.567982	0.753646	0.6540	1.025965	*	*
				1.594174	0.382202	0.874711	0.935260	2.4470	1.262606	0.804470	1.358
				3.276329	0.093121	0.985816	0.992883	10.6623	2.876861	0.990107	9.409
				1.133635	0.016936	1.012252	1.006107	59.4064	1.064723	1.004348	6.431
8	0.1	35	0.0364	1.083222	1.084425	0.488806	0.699147	0.6447	1.040780	*	*
				1.759678	0.422272	0.822698	0.907024	2.1480	1.326529	0.773513	1.370
				9.760872	0.118048	0.974679	0.987258	8.3682	3.124239	0.982941	7.548

LP 880Hz	1-stage	2-stage	3-stage
C1 and C2	10nF + 6·8nF parallel	10nF + 6·8nF parallel	10nF only
R1	390kΩ	820kΩ	910kΩ
R2, R3, R7	20kΩ	13kΩ	18kΩ
R4	43kΩ	82kΩ	470kΩ
R5	91k + 2·7kΩ in series	100kΩ + 5·1kΩ in series	39kΩ
R6 and R8	10kΩ	10k	20kΩ

First-order stage C1 = 10nF and R1 = 120kΩ.

HP 760Hz	1-stage	2-stage	3-stage
C1 and C2	10nF	10nF	10nF
R1 and R4	24kΩ	110kΩ	510kΩ
R2	110kΩ	180kΩ	39kΩ
R3 and R7	11kΩ	18kΩ	20kΩ + 240Ω in series
R5, R6, R8	20kΩ	20kΩ	20kΩ

First-order stage C1 = 1·2nF and R1 = 27kΩ.

HP 820Hz	1-stage	2-stage	3-stage
C1 and C2	10nF	10nF	10nF
R1 and R4	22kΩ	110kΩ	470kΩ + 20kΩ in series
R2	100kΩ	100kΩ + 68kΩ in series	36kΩ + 1kΩ in series
R3 and R7	10kΩ	16kΩ	18kΩ + 1kΩ in series
R5, R6, R8	20kΩ	20kΩ	20kΩ

First-order stage C1 = 1·2nF and R1 = 24kΩ.
All capacitors were of five per cent tolerance, and resistors one per cent.

Fig 12 shows the block diagram of a seventh-order lowpass filter, where one first-order stage is followed by three second-order stages in cascade. The input resistor shown is only necessary if no dc return to ground is provided, eg in case of capacitive coupling to a first-order lowpass stage, then a value up to 1kΩ should be inserted. Under conditions where the input is taken direct from a low impedance speaker source to ground, the input resistor can be deleted. The filter output is capable of driving high-impedance loads direct (ie >600Ω headphones), but should have an additional audio amplifier to match low-impedance speaker loads. An LM380 requires the least external components, but will contribute to increase the noise floor.

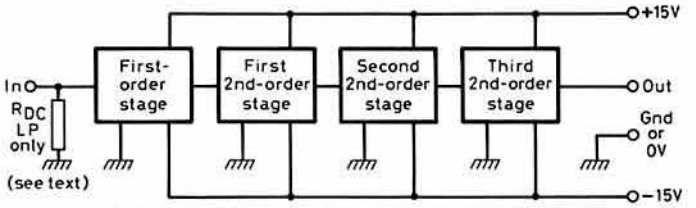

Fig 12. Block diagram, seventh-order filter stages

Fig 13 shows the block diagram of the cw filter in its entirety, and Fig 14 indicates where each of the 12 second-order stages, plus the four single-order stages are located.

The construction of the filters is less critical, since rf is not directly involved. However, for best results the filters should be housed in a small diecast box, measuring 120 by 95 by 34mm (source: Farnell), which allows a mounted pcb, toggle switches and sockets, plus optional extras to be fitted.

The mounting of components could be carried out on "breadboard" or matrix board, where connections are made using component leads.

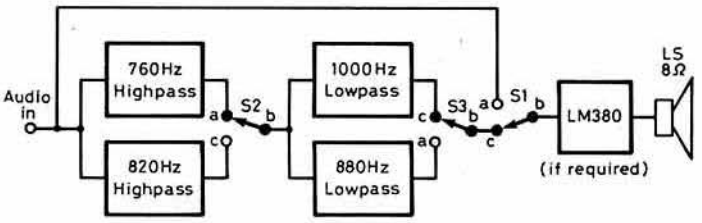

Fig 13. Block diagram, cw bandpass filter

Table 4. G0CKZ's computerized biquad elliptic lowpass active filter design summary

To design a seventh-order biquad elliptic **lowpass** filter, select standard value resistors and capacitors as close as possible to the calculated values and construct the filter, or its stages, in accordance with Fig 1(a) and Fig 1(b).

fc = 1,000Hz
fs = 1,361Hz
TW = 361Hz

N = 7
PRW = 3·00dB
MSL = 80·0dB
K = 1 or 0dB

Coefficients used

A	B	C
2·646568	0·244509	0·275048
7·328972	0·130851	0·716553
1·915736	0·038211	0·977841
		0·152661

First-order stage results (see Fig 1(b))
C1 = 10·000nF
R1 = 104·25kΩ
fp = 391Hz

Second-order stage results (see Fig 1(a))

	1-stage	2-stage	3-stage
Qp	2.14	6·47	25·9
fp	524Hz	846Hz	989Hz
fz	1,627Hz	2,707Hz	1,384Hz
fm	489Hz	840Hz	988Hz
Km	6·04dB	15·4dB	22·1dB
Km	2·00	5·86	12·7
C1	10·000nF	10·000nF	10·000nF
C2	10·000nF	10·000nF	10·000nF
R1	626·33kΩ	1·24MΩ	816·02kΩ
R2	30·35kΩ	18·80kΩ	16·09kΩ
R3	30·35kΩ	18·80kΩ	16·09kΩ
R4	65·09kΩ	121·63kΩ	416·52kΩ
R5	192·44kΩ	204·56kΩ	39·18kΩ
R6	20·00kΩ	20·00kΩ	20·00kΩ
R7	30·35kΩ	18·80kΩ	16·09kΩ
R8	20·00kΩ	20·00kΩ	20·00kΩ

Comments
(a) The performance of the filter remains unchanged if all the resistors are multiplied and all the capacitors are divided by a common factor.
(b) The input impedance of the op-amp should be at least 12·44MΩ; its open loop gain should be at least 50 times the amplitude of the filter at fc = 1,000Hz; and the desired peak-to-peak voltage at fc should not exceed 1E6/pi*fc times the slew-rate of the op-amp. A dc return path to ground should be provided at the filter input.
(c) Tuning of the circuit may be accomplished by adjusting R5 to set the stage zero frequency at fz, adjusting R3 to set the stage peak frequency fm, adjusting R4 to set the stage pole-pair quality factor Qp, and adjusting R1 or R2 to set the stage gain Km.

Table 5. G0CKZ's computerized biquad elliptic highpass active filter design summary

To design a seventh-order biquad elliptic **highpass** filter, select standard value resistors and capacitors as close as possible to the calculated values and construct the filter, or its stages, in accordance with Fig 1(a) and Fig 1(c).

fc = 760Hz
fs = 559Hz
TW = 201Hz

N = 7
PRW = 3·00dB
MSL = 80·0dB
K = 1 or 0dB

Coefficients used

A	B	C
2·646568	0·244509	0·275048
7·328972	0·130851	0·716553
1·915736	0·038211	0·977841
		0·152661

First-order stage results (see Fig 1(c))
C1 = 1·200nF
R1 = 26·64kΩ
fp = 1,945Hz

Second-order stage results (see Fig 1(a))

	1-stage	2-stage	3-stage
Qp	2·14	6·47	25·9
fp	1,449Hz	898Hz	769Hz
fz	467Hz	281Hz	549Hz
fm	1,556Hz	904Hz	769Hz
Km	6·04dB	15·4dB	22·1dB
Km	2·00	5·86	12·7
C1	10·000nF	10·000nF	10·000nF
C2	10·000nF	10·000nF	10·000nF
R1	23·56kΩ	114·68kΩ	535·90kΩ
R2	105·68kΩ	181·31kΩ	40·57kΩ
R3	10·98kΩ	17·73kΩ	20·71kΩ
R4	23·56kΩ	114·68kΩ	535·90kΩ
R5	20·00kΩ	20·00kΩ	20·00kΩ
R6	20·00kΩ	20·00kΩ	20·00kΩ
R7	10·98kΩ	17·73kΩ	20·71kΩ
R8	20·00kΩ	20·00kΩ	20·00kΩ

Comments
(a) The performance of the filter remains unchanged if all the resistors are multiplied and all the capacitors are divided by a common factor.
(b) The input impedance of the op-amp should be at least 5·36MΩ; its open loop gain should be at least 50 times the amplitude of the filter at fc = 760Hz, and the desired peak-to-peak voltage at fc should not exceed 1E6/pi*fc times the slew-rate of the op-amp. A dc return path to ground should be provided at the filter input.
(c) Tuning of the circuit may be accomplished by adjusting R5 to set the stage zero frequency at fz, adjusting R3 to set the stage peak frequency fm, adjusting R4 to set the stage pole-pair quality factor Qp, and adjusting R1 or R2 to set the stage gain Km.

1 LP stage 880Hz plus single-order	3 HP stage 820Hz	2 HP stage 820Hz	1 HP stage 820Hz plus single-order
2 LP stage 880Hz	1 HP stage 760Hz plus single-order	2 HP stage 760Hz	3 HP stage 760Hz
3 LP stage 880Hz	3 LP stage 1000Hz	2 LP stage 1000Hz	1 LP stage 1000Hz plus single-order

(viewed from track side)

Fig 14. Individual filter stage locations

Veroboard with copper tracks running in one direction is another choice, but tracks will have to be cut in order to prevent short circuits. Fig 15 shows the track side of a pcb layout in double size, where a total of 12 second-order sections, or 12 ic locations are marked (three rows of four ICs each). The centre row runs opposite to the outer ones, and care should be taken when mounting components on to the pcb. The accompanying component layout can be seen in Fig 16, where links are marked to indicate the untracked op-amp pin 11, going to its negative rail. Ground or 0V is the only rail connected to all appropriate ic pins, and both the positive and negative rails will have to be bridged across the tracks with a length of insulated wire. IC sockets are recommended but not essential, to save cost. Care should be taken to position each ic correctly, since reverse polarity rails will lead to the destruction of four op-amps. After building the first section, one quickly becomes familiar with the process and it becomes routine to copy the remaining eleven stages. Increasing the value of a resistor to obtain, for example, the odd 416kΩ needed in the third 1kHz lowpass second-order stage, put a 390kΩ and 27kΩ preferred-value resistor in series, by inserting both into the location of R4, joining them together above the pcb with a spot of solder. To decrease a resistor value for tuning purposes, put the second resistor on the track side, in parallel with the original one.

At least two additional components are required to build a first-order stage. The 1kHz lowpass, for example, uses the three vacant ic pins 8, 9 and 10, where its first second-order stage is located. There, a 10nF capacitor is fitted to the track side, going from pin 10 to the centre rail of 0V. The 100kΩ filter input resistor is left standing on the component side, one end only connected, to ic pin 10. A jumper wire links pins 8 and 9.

Again, bare jumper wires are used on the component side to link the filter input and output stages. Here, some cut-off resistor lead-end pieces can be

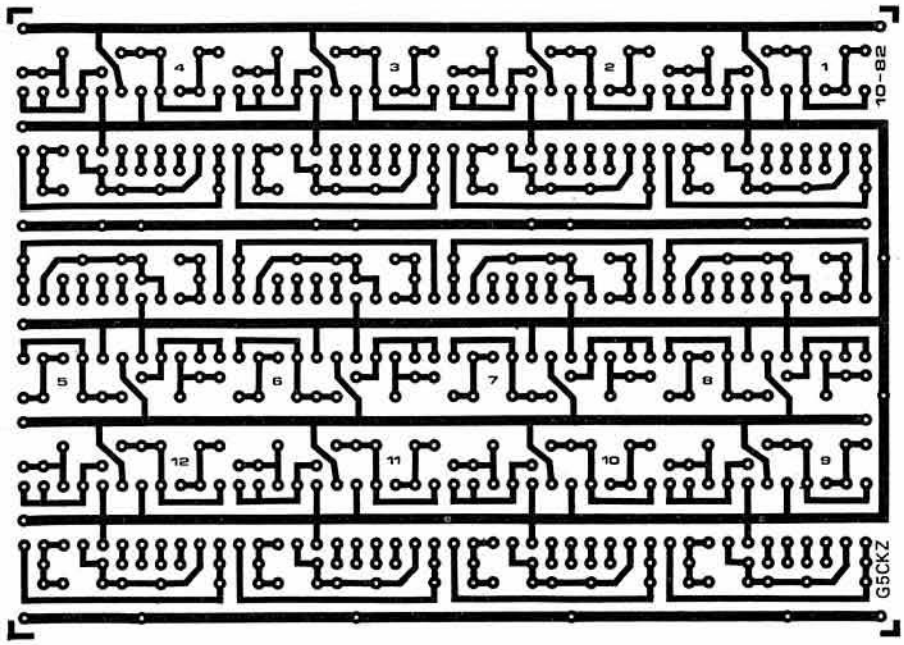

Fig 15. PCB track layout

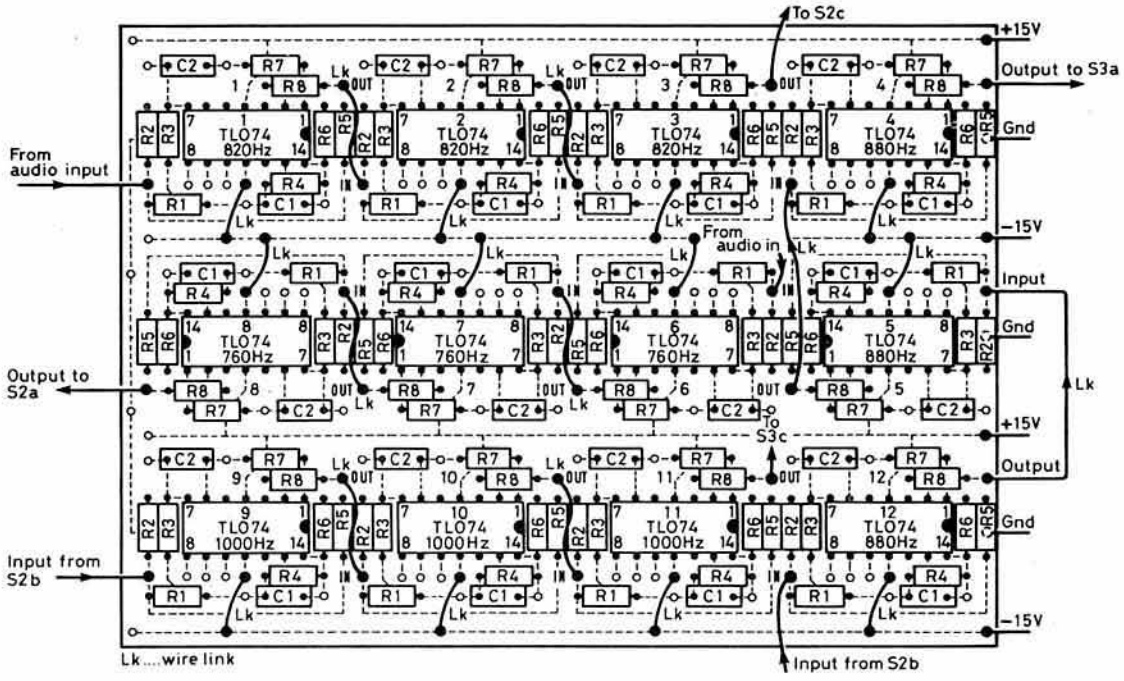

Fig. 16. PCB component layout

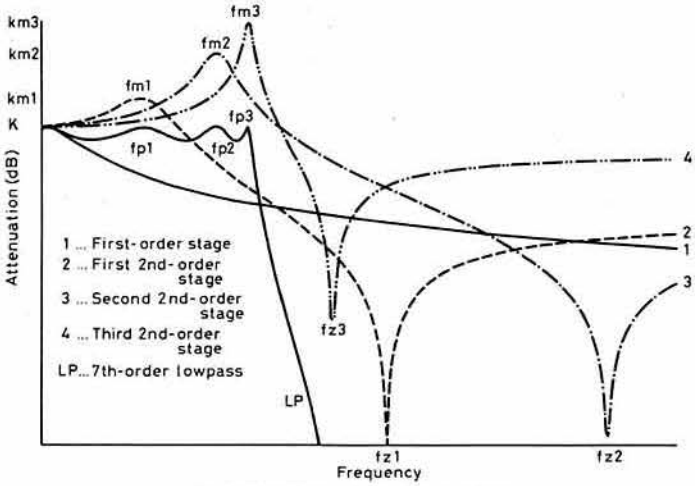

Fig 17. Lowpass tuning responses

put to good use by soldering one of them into each of the holes marked i/p and o/p, then temporarily joining the ends above board for easier accessibility when the circuit function is being verified, or to isolate a stage for tuning purposes. One pair is needed to join the o/p from the first-order stage to the i/p of the first second-order stage, a second pair to link the o/p of the first stage to the i/p of the second stage, and finally a third pair from the o/p of the second stage to the i/p of the third.

For cmos or ttl it is good design practice to tie unused inputs to one rail in order to prevent unforeseen problems. The same can be said for op-amp design. The unused pins 8 and 9 of spare op-amps should be linked together, and the non-inverting input (pin 10) taken to ground. Apart from first-order stages (see Figs 1(b) and 1(c)), the vacant op-amp pins can be utilized for other circuit functions. A rail divider is shown in Fig 1(d), where the ratio of the two resistors on the non-inverting input determines the output voltage from a low-impedance source. If these resistors are made equal in value, say 100kΩ each, then the resulting output voltage will be half of the voltage applied. That is, if the filter were to be used on a single + 12V supply (ie car battery), it could provide the reference voltage for all non-inverting inputs, rather than have them go to ground. Care would have to be taken in coupling the filter input and output sections, because of a plus 6V dc level shift.

Other applications may include a summing amplifier or a buffer circuit like those in Figs 1(e) and 1(f). The summing amp for example, would produce a notch filter if its inputs were connected to the outputs of a highpass and a lowpass.

Circuit details of a power supply have been omitted, since voltage variations have little or no effect on the performance of the filter. Any dual-power supply providing a positive and a negative rail from as low as 3V, and as high as 18V, can be used. Voltages in the range of 12V to 15V are recommended for best dynamic performance. The current consumption of a filter using TL074s or TL084s may be estimated by allowing a maximum supply current of about 2·5mA per op-amp. In our case the dual supply should be capable of delivering 4*12*2·5mA = 120mA total current.

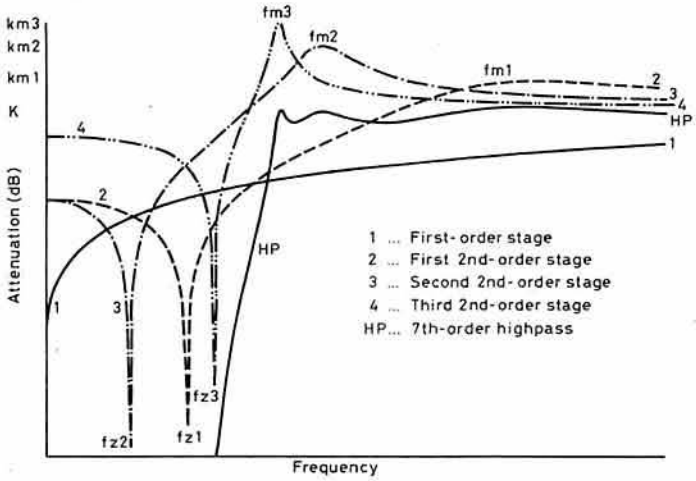

Fig 18. Highpass tuning responses

The effect of cascading is best demonstrated graphically. For example, Fig 17 shows the seventh-order lowpass tuning responses, where each stage exhibits its own individual, non-interacting characteristic and contributes to the overall desired lowpass transfer function. Each second-order stage has its own specific peak frequency (fm1, fm2, fm3) at the level (km1, km2, km3), and zero frequency (fz1, fz2, fz3) to make up the overall lowpass response with its pole-pair frequencies (fp1, fp2, fp3) at the level K.

The first-order stage only has a shallow monotonically-descending slope, starting at the reference level K. Even-order filters (N = 2,4,6 . . .) produce a starting level, which is equal to the amount of passband ripple subtracted from K.

Fig 18 shows the individual and overall responses for a seventh-order highpass. Tuning is best accomplished by isolating a particular second-order stage and measuring its zero frequency. This can be done using an audio generator and a sensitive oscilloscope by adjusting the generator for minimum amplitude display. If the generator accuracy is doubtful, then its output should be measured with a frequency counter. The aim then is to adjust the resistor marked R5 until the calculated zero frequency agrees with the measured generator frequency.

If at all necessary, zero frequency tuning will suffice in this particular design. Additional trimming is possible where needed, ie in close tolerance filters (PRW = 0·1dB).

More tuning details are provided in Tables 4 and 5, under (c) in comments.

Measured results

All results were taken with a network analyzer, allowing for two input measurements in the frequency domain, ie phase and group delay measurements. Hardcopy of the enclosed results was obtained from a video plotter, directly interfaced to the analyzer display. These plots confirm that the designed cw bandpass filters really do live up to the standard of their specifications. The plotted lowpass filter ripples in Fig 4 occupy the top third of the plot, where the frequency starts to scan as indicated at 10Hz (left), and ends 90s after at 1,010Hz (right). The vertical scale for the passband ripple is calibrated in 1dB per major division, and we notice the maximum deviation being three major divisions of 3dB, with its cut-off frequency at fc = 998Hz. The stopband ripple is superimposed into the same display window, but a different decibel and frequency range apply, where each major division on the vertical axis now becomes 5dB, totalling a stopband ripple width of greater than 30dB. The frequencies scanned cover the range from 1·3kHz to 3kHz; 2kHz of the lowpass filter shape being displayed on a linear frequency scale in Fig 3. The measured transition width between the two points marked with an asterisk is TW = 360Hz, and the minimum stopband loss is MSL = 80dB.

Fig 19 shows the sort of response one is likely to find in sales literature, because the slopes appear to be steeper than those in Fig 8, but indeed are identical. Here the analyzer start frequency is 100Hz and stop frequency 10kHz, producing a dissimilar logarithmic scale, and displaying perfect symmetrical slopes. The highpass responses are displayed in Figs 6 and 7,

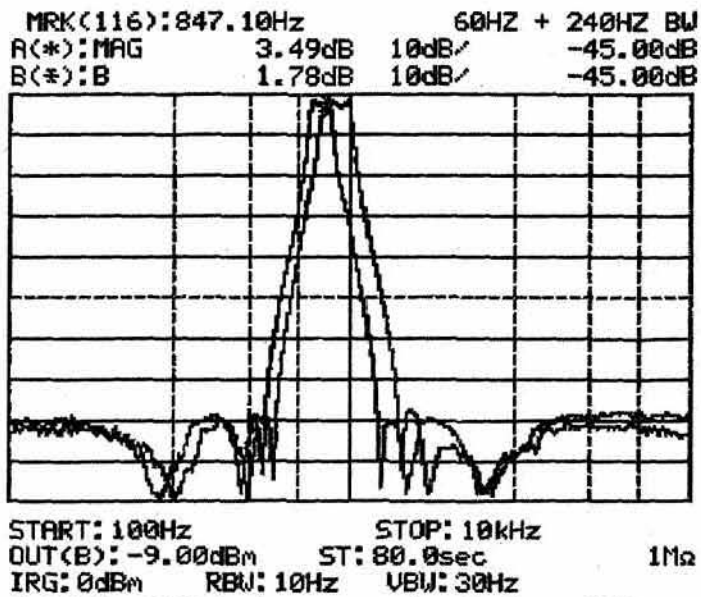

Fig 19. 60Hz and 240Hz bandpass responses (logarithmic), cascaded lowpass and highpass filters. N = 7, prw = 3dB, msl = 80dB

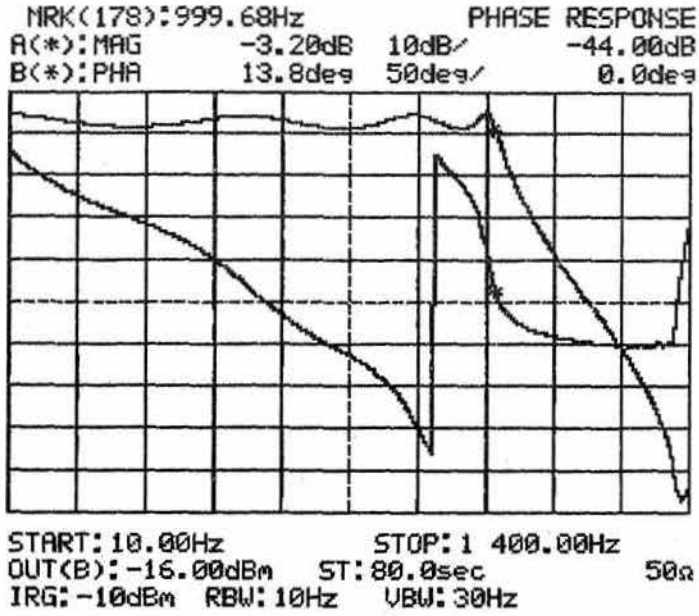

Fig 20. 1kHz lowpass phase response

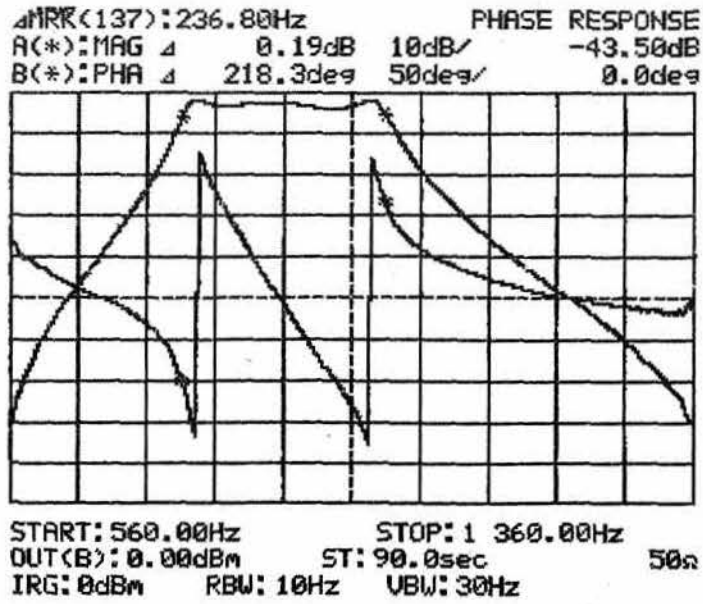

Fig 22. 240Hz bandpass phase response

and the bandpass in Figs 8 and 9. Special responses not normally found in general filter specifications are displayed in Figs 20 and 22, for the phase response inside the pass- and transition bands. Delay responses are shown in Fig 21 for the 1kHz lowpass, and Fig 23 for the 240Hz wide bandpass where a maximum delay coincides with the filter stages of maximum gain, near fc. Linearity becomes a prime requirement for phase in allpass filters and constant delays in time-delay filters, and are of less importance in frequency selective filter design. As a final set of the measured bandpass performance data, find the actual shape factors for all four possible combinations:

			Shape factors		
	Range (Hz)	Bandwidth	30/3dB	60/6dB	80/6dB
BP1	820–880	60Hz	4·2	6·6	7·8
BP2	760–880	120Hz	2·4	3·9	4·7
BP3	820–1,000	180Hz	1·9	3·2	3·9
BP4	760–1,000	240Hz	1·7	2·8	3·2

Conclusion

It is my intention to make available the necessary data for active elliptic audio filter design. Aiming to stimulate home construction for the novice and expert alike, by shedding some light on obscure theories and making it user-orientated, by keeping its maths to an absolute minimum.

The mathematical involvement could deter some constructors, but tedious calculations are quickly solved by calculator or home computer. Formulas in the previous step-by-step calculations were written in such a fashion that they could easily be adapted as "one-liners" in any Basic language.

The active filters described can be designed for any audio frequency and, indeed, well above if faster op-amps are used. Fixed-frequency filters lend themselves to applications like cw and rtty, plus many more, where only the receiver main tuning dial determines the signal to which we are listening. Here, a combination of lowpass and highpass elliptic filters were used to create a bandpass response, as was practically demonstrated.

The total hardware and component cost incurred to build this high performance cw filter has been estimated to reach about £40.

Should demand justify the etching of a universal pcb, I will try to make one available. Any requests or suggestions on this subject matter are welcome—please send saes.

References

[1] "A cw filter for the radio amateur newcomer", Edward E Wetherhold, W3NQN. *Rad Com* January 1985.

[2] "Elliptic lowpass audio filter design using miniature preferred value components", S Niewiadomski, BRS54049. *Rad Com* October 1984.

[3] "Simplified elliptic lowpass filter construction using surplus 88mH inductors", Edward E Wetherhold, W3NQN. *Rad Com* April 1983.

(Continued on page 189)

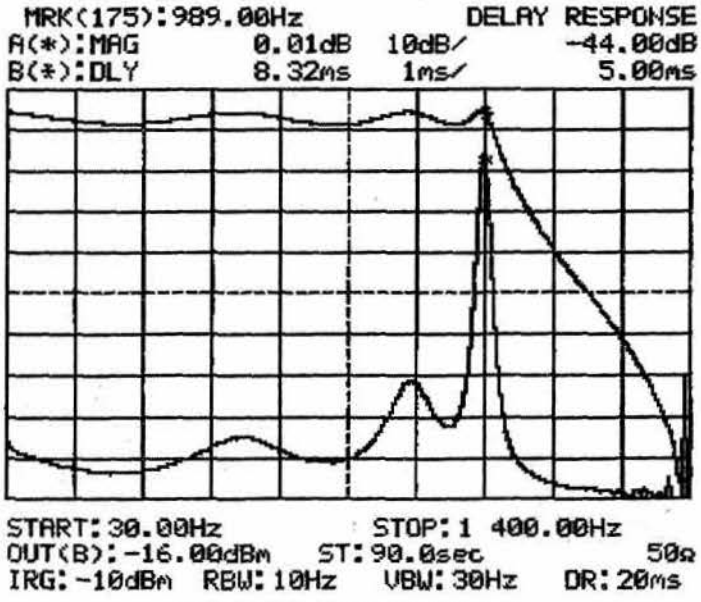

Fig 21. 1kHz lowpass delay response

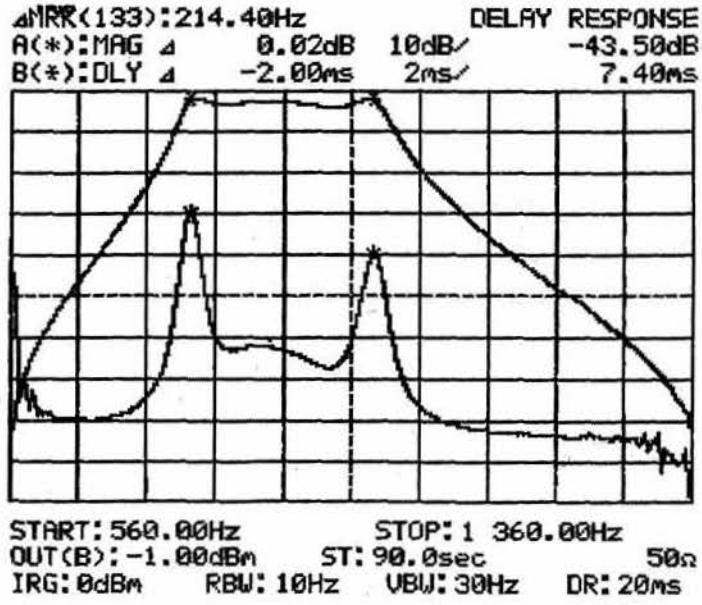

Fig 23. 240Hz bandpass delay response

On the other hand there are already reports of quite severe interference to reception between 3·5 and 3·6MHz from harmonics of 1·6 to 1·8MHz base transmitters. Some amateurs are already reporting difficulty in using the cw end of 3·5MHz, although it is uncertain how much of such interference results from legal units and how much to models not conforming to the DTI specification. But it would hardly be surprising if any low-cost local 1·7MHz transmitter radiates sufficient second harmonic to cause interference on the 3·5MHz band, and quite possibly spurious signals at other frequencies: 1·750 to 1·800MHz, unfortunately, is harmonically related to the 3·5, 7, 14, 21 and 28MHz amateur bands!

I have no information on whether the RIS or BT are preparing to take any action to recognize and/or minimize this problem, which is still largely confined to a few areas. It seems odd that telephone users who react so strongly to any suggestion that their telephones may be "tapped" seem to be quite willing to let their private conversations be picked up (with slope detection in the absence of an fm discriminator) at a considerable distance! Perhaps they have no idea how far their words are winging! Even a few milliwatts do not stay within a user's own estate! Perhaps if they were made more aware of the facts of radio there would be far less than the millions of users anticipated by the year 2000, at least on the 1·7/49MHz bands.

Multiway 6A adapters

While some of us still live in houses having the old round three-pin 15A, 5A, 2A mains wall sockets, the majority must by now be using the standard flat-pin 13A socket. This often implies the use in the shack of a quite large four- or six-way junction box, or at least the rather cumbersome three-way 13A adapter.

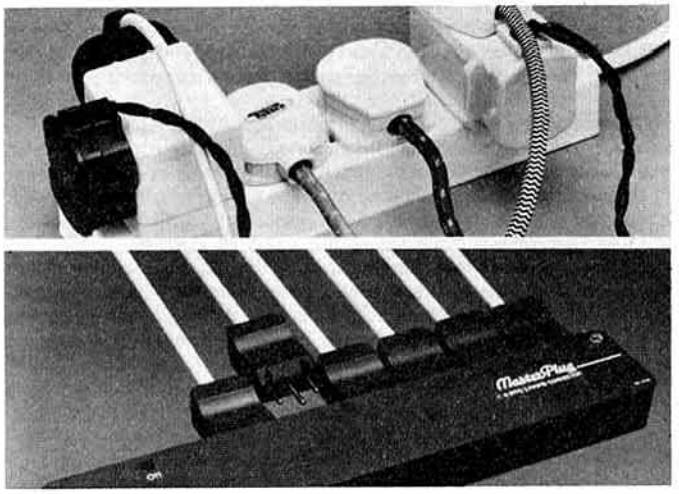

Photo 1. (Top) Conventional six-way junction box adapter with 13A sockets. (Below) The Conblock six-way adapter

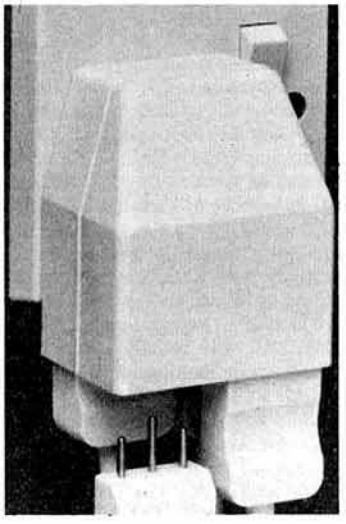

Photo 2. The new Conblock plug-adapter providing four outlets with 6A sockets

A recent press release, issued on behalf of Conblock Electrical Ltd (0686-27100) of Newtown, Powys, sings the praises of two new adapters, sold complete with plugs. They provide compact three-pin 6A sockets, so making for much smaller and lighter units for the many applications where the individual loads do not really require the use of 13A plugs/sockets (ie loads of less than about 1kW), and particularly where there are several separate units.

Conblock claim the new adapters are available at most electrical high street outlets and confirm fully with BS5733 and the latest electrical equipment safety regulations.

An in-line adapter for use with 13A sockets provides six in-line 6A shuttered sockets. They are also making a wallplug adapter that plugs into a 13A socket and provides four of the 6A sockets.

That earphone filter!

The January *TT* included an item on putting a resonant reed as the effective part of the diaphragms of a high-impedance as described in the Australian *Radio Experimenter,* of unknown date. I expressed reservations about this idea, particularly the "lag". It all seemed a little too good to be true. What I did not realize, until reminded by Eric Sandys, GI2FHN, was that exactly the same idea had been previously published in *QST*, but with a few significant differences! It was in the *April* (1976) issue and was ascribed to "Tin Ehres, WH0OP". Quite clearly it was one of the traditional semi-believable *QST* April Fool jokes!

My apologies to readers for including this idea in a January rather than an April issue—and thanks to GI2FHN.

There is, of course, no fundamental reason why the output transducer or impedance matching transformer should not constitute an effective filter. Douglas Byrne, G3KPO, writes:

"A horn loudspeaker from the 'twenties makes an excellent af filter for cw. Its 800Hz peak can be much improved by bunging a large wad of cotton-wool right down the aperture! Most of these early loudspeakers had high-impedance windings (about 2,000Ω) but can be fed from low-Z sockets via almost any small output transformer wired backwards (ie with the low-Z winding connected to the receiver)."

The German "diplomatic" suitcase set (*TT* September 1985) used the technique of resonating the output transformer, and in a pre-war *T & R Bulletin* an article showed how by cutting a gap in the core, the *Q* could be improved and the output peaked at the required frequency. □

ACTIVE ELIPTIC AUDIO FILTER DESIGN USING OP-AMPS

(Continued from page 184)

[4] "An introduction to elliptic filters for the radio amateur", John Wilkinson, G4HGT. *Rad Com* February 1983.

[5] *A Handbook of Active Filters*, D E Johnson, J R Johnson, H P Moore. Prentice Hall 1980.

[6] *Synthesis of Passive Networks*, E A Guillemin. John Wiley & Sons 1957.

[7] *Manual of Active Filter Design*, J L Hilburn, D E Johnson. McGraw-Hill 1973.

[8] *Active Filter Cookbook*, Don Lancaster. Howard W Sams & Co 1979.

[9] *Solid State Design for the Radio Amateur*, W Hayward, D DeMaw. ARRL 1977.

[10] *Operational Amplifiers—Design and Applications*, J G Graeme, G E Tobey, L P Huelsman. McGraw-Hill 1971.

APPENDIX

In order that the author can process your personal filtering requirements and to generate the calculated printouts as mentioned in this article, a suggested format for specifying these requirements should be adopted. Find below the described 1,000Hz lowpass as an example:

CUSTOM DESIGNED ACTIVE ELLIPTIC AUDIO FILTERS

Specify your filter requirements as follows:

(1) Tick filter type: LOWPASS HIGHPASS BANDPASS NOTCH				
(2) Passband frequency:		fp	1,000	Hz
	or	fp1	.....	Hz
	and	fp2	.....	Hz
(3) Stopband frequency:		fs	1,361	Hz
	or	fs1	.....	Hz
	and	fs2	.....	Hz
(4) Maximum allowable passband ripple: (select 0·1, 0·5, 1, 2 or 3dB)		PRW	3	dB
(5) Minimum stopband loss: (between 30dB and 100dB)		MSL	80	dB
	or	MSL1	.....	dB
	and	MSL2	.....	dB
(6) Maximum allowable filter gain or loss:		K	0	dB
	or	K	1	
(7) Source impedance: (if known)		Rs	600	Ω
(8) Load impedance:		Rl	600	Ω
(9) Maximum input signal level:		Vin	3	Vpp
(10) Available supply voltages: (usually dual rails, between 3V and 18V)		Vdc	+/−15	V

(11) Special points to consider:.....
eg low Qp, limited complexity of the filter, restricted size, close tolerance in attenuation or ripple values, critical phase or delay characteristics in the passband, DC drifts and offsets.

□